ACEEE Proc. of Int. Conf. on Recent Trends in Information, Telecommunication and Computing, ITC

Effectiveness of Optimal Preamble based Channel Estimation for FBMC/OQAM

V. SandeepKumar [1] and Dr.S.Anuradha [2]
[1] National Institute of Technology Warangal/ECE, Warangal, India
Email: sandeepvangala443@gmail.com
[2] National Institute of Technology Warangal/ECE, Warangal, India
Email: anuradha@nitw.ac.in

***Abstract*— Filter Bank-based Multicarrier Modulation (FBMC) using Offset Quadrature Amplitude Modulation (OQAM), known as FBMC/OQAM (or OFDM/OQAM), provides an attractive alternative to the conventional Cyclic Prefix-Based Orthogonal Frequency Division Multiplexing (CP-OFDM), especially in terms of increased robustness to frequency offset and Doppler spread, and high bandwidth efficiency. It suffers, however, from an inherent (intrinsic) imaginary Intercarrier/Inter-symbol interference that complicates signal processing tasks such as Channel Estimation (CE). Recently, the so-called Interference Approximation Method (IAM) was proposed for preamble-based CE. It relies on the knowledge of the pilot's neighbourhood to approximate this interference and constructively exploit it in simplifying CE and improving its performance. The IAM preamble with nulls at the neigh boring time instants and extended version of it, which can provide significant improvement through an appropriate exploitation of the interfering symbols from neighbouring time instants that results in CE performance was recently reported. In this paper, we present IAM preamble design and apply it to identify the optimal IAM preamble sequence which results in a higher gain. Numerical results have verified the effectiveness of the theoretical framework and a gain of 1.24 dB against E-IAM-C.**

***Index Terms*— FBMC/OQAM, CE (Channel Estimation), IAM (Interference Approximation Method), E-IAM-C (Extended IAM-C), Pilots.**

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) has become quite well-liked in each wired and wireless communications [4, 2], in the main owing to its immunity to multipath attenuation, which allows a significant increase in the transmission rate [13].using the cyclic prefix (CP) as a guard interval, OFDM manages to show a frequency selective channel into a set of parallel flat channels with independent noises. This greatly simplifies each the estimation of the channel and the recovery of the transmitted data at the receiver. However, these advantages come at the cost of an enhanced sensitivity to frequency offset and Doppler spread. This can be as a result of the fact that, although the subcarrier functions are perfectly localized in time, they suffer from spectral leak in the frequency domain and hence inter-carrier interference (ICI) results. Moreover, the inclusion of the CP entails a waste in transmitted power further as in spectral efficiency, which, in practical systems, can go up to twenty five percent [2].

Filter bank-based multicarrier modulation (FBMC) using offset quadrature AM (OQAM), known as FBMC/OQAM or OFDM/OQAM [8], provides an alternative to CP-OFDM that can mitigate these

DOI: 02.ITC.2014.5.559

drawbacks. FBMC/OQAM employs pulse shaping via an IFFT/FFT based efficient filter bank, and staggered OQAM symbols, i.e., real symbols at doubly the symbol rate of OFDM/QAM, are loaded on the subcarriers [12]. This allows for the pulses to be well localized in each the time and the frequency domains while still keeping most spectral efficiency. As a consequence, the system's robustness to frequency offsets and Doppler effects is enhanced and at the same time an enhanced spectral containment, for information measure sensitive applications (e.g., cognitive radio [1]), is offered. Moreover, FBMC/OQAM doesn't need the inclusion of a CP, which may result in even higher transmission rates [12].

However, the subcarrier functions are now only orthogonal in the real field, which means that there's perpetually an intrinsic imaginary interference among (neighboring) subcarriers and symbols [7]. As a consequence, the simplicity of the channel estimation (CE) in CP-OFDM is lost in FBMC/OQAM systems, because the real-valued pilots also are polluted" by this imaginary-valued contribution of their neighbours. A number of training schemes and associated channel estimation methods that take this under consideration have recently appeared in the literature, together with each preamble-based (e.g., [9,11]) and scattered pilots-based (e.g., [7,10,3]) ones. These can be roughly characterised as aiming at cancelling the unwanted interference or constructively exploiting it to boost the estimation performance. The latter approach depends on the fact that the interference to a pilot symbol is principally contributed to by its first-order neighbors, and hence it may be approximated if these neighbours also carry known symbols, a not uncommon situation within a preamble. This has been known as the Interference Approximation method (IAM). The idea is to combine the real-valued pilot with the interference approximation into a complex-valued pseudo-pilot and performing channel estimation as in CP-OFDM, beneath the idea of a locally constant channel frequency response. As the pseudo pilot divides the noise components, its magnitude should be as large as possible. This way, the interference turns into a benefit for the channel estimation task and hence the neighboring pilots should be therefore chosen on maximize its effect.

A number of variants of IAM have been suggested, corresponding to different alternatives for the preamble structure [18]. Those that have obtained most of the attention comprise three FBMC/OQAM symbols (i.e., 1.5 complex OFDM/QAM symbols), with the edge ones being all zeros. This is often to safeguard the center vector of pilots from being interfered by the unknown knowledge components of the previous and current frames the center pilot symbols are then chosen therefore on maximize the interference contributions from neighboring subcarriers. Recently, such a preamble structure with imaginary pilots was suggested in [16] and individually in [6] and shown to be the optimum alternative inside this kind of preambles. It was called IAM-C to signify the presence of complex (imaginary) pilots, in analogy with IAM-R [9], the optimum real-valued preamble. IAM-C advances upon IAM-I [11], a preceding IAM variant furthermore employing invented pilots, by maximizing the magnitude of all pseudo-pilots rather than of only one third of them as in IAM-I.

The possible profits from a preamble that has nonzero pilots at its side FBMC/OQAM symbols as well. Recently, a preamble structure [17] of this kind is proposed that exploits the symmetry of the interference weights and their relation values to get pseudo-pilots of a magnitude larger than in IAM-C. Recently, the expands IAM-C [17] to incorporate edge symbols as well is proposed, it will be called expanded IAM-C (E-IAM-C).

Motivated by the standard of IAM-C method and awareness of the nature of the three IAM methods in [17, 11], in this paper we have formulated a general theoretical structure for IAM preamble design. As a outcome the optimal preamble IAM sequence has been recognised which results in a higher performance gain. The effectiveness of the theoretical framework and the superiority of proposed preamble has thereafter been verified by numerical simulation.

The rest of the paper is organized as pursues: In part II, the discrete-time baseband equivalent model for the FBMC/OQAM scheme is recounted. Part III reconsiders the three major preamble structures that encompass edge none guard symbols, IAM-R, IAM-C and E-IAM-C. The suggested optimal preamble is developed and considered in part IV. Replication outcomes are described in part V. Part VI concludes the paper.

II. SYSTEM MODEL

The (QAM or OQAM based) OFDM modulator output is transmitted through a channel of length L_h, which is assumed to be invariant in the duration of the preamble. At the receiver front-end, noise is added, which is

assumed Gaussian with zero mean and variance σ^2 . For CP-OFDM, the classical Configuration is assumed [13]. The discrete-time signal at the output of an FBMC/OQAM synthesis filter bank (SFB) is given by [12]

$$s(l) = \sum_{m=0}^{M-1} \sum_{n} d_{m,n} g_{m,n}(l), \tag{1}$$

Where $d_{m,n}$ are real OQAM symbols, and $g_{m,n}(l) = g\left(l - n\frac{M}{2}\right) e^{j\frac{2\pi}{M}m\left(l - \frac{L_g - 1}{2}\right)} e^{j\varphi_{m,n}}$, with g being the real symmetric prototype filter impulse response (assumed here of unit energy) of length L_g, M being the even number of subcarriers, and $\varphi_{m,n} = \varphi_0 + \frac{\pi}{2}(m+n) \bmod \pi$, where φ_0 can be arbitrarily chosen [12]. In this paper, and without loss of generality, $\varphi_{m,n}$ is defined as $(m+n)\frac{\pi}{2} - mn\pi$ as in [12]. The filter g is usually designed to have length $L_g = KM$,with K being the overlapping factor. Here the transmitted symbols $d_{m,n}$ are real-valued with symbol duration τ_0 and inter carrier spacing ν_0 respectively. Two kinds of realisations of OFDM/OQAM are of practical interest. One can either set $\nu_0 = F$, $\tau_0 = T/2$ as in [14], [15] or set $\tau_0 = T$, $\nu_0 = F/2$ as in [14]. We use the former approach. The double subscript $(.)_{m,n}$ denotes the $(m,n)^{\text{th}}$ frequency time (FT) point. Thus, m is the subcarrier index and n the OQAM symbol time index. The pulse g is designed so that the associated subcarrier functions $g_{m,n}$ are orthogonal in the real field [12]. This implies that even in the absence of channel distortion and noise, and with perfect time and frequency synchronization, there will be some inter carrier (and/or intersymbol) interference at the output of the Analysis Filter Bank (AFB), which is purely imaginary (the notation in [9] is adopted):

$$\sum_{l} g_{m,n}(l) g^{*}_{p,q}(l) = j\langle g\rangle_{m,n}^{p,q}, \tag{2}$$

and known as intrinsic interference [7]. Making the common assumption that the channel is (approximately) frequency flat at each subcarrier and constant over the duration of the prototype filter [9], which is true for practical values of L_h and L_g and for well time-localized g's, one can express the AFB output at the p$^{\text{th}}$ subcarrier and q$^{\text{th}}$ FBMC/OQAM symbol as:

$$y_{p,q} = H_{p,q} d_{p,q} + \underbrace{j \sum_{m=0}^{M-1} \sum_{n(m,n)\neq(p,q)} H_{m,n} d_{m,n} \langle g\rangle_{m,n}^{p,q}}_{I_{p,q}} + \eta_{p,q} \tag{3}$$

Where $H_{p,q}$ is the Channel Frequency Response (CFR) at that FT point, and $I_{p,q}$ and $\eta_{p,q}$ are the associated interference and noise components, respectively. $\eta_{p,q}$ has been shown to be stationary (in fact, also Gaussian with zero mean and Variance σ^2) and correlated among adjacent subcarriers (see [5]). This correlation will be henceforth assumed negligible, a valid assumption for well frequency-localized prototype filters.

A common assumption is that, with a well time frequency localized pulse, contributions to $I_{p,q}$ only come from the first-order neighbourhood of (p,q) , namely $\Omega_{p,q} = \{(p\pm1,q\pm1),(p,q\pm1),(p\pm1,q)\}$. Moreover, the CFR is almost constant over this neighborhood, one can write (3) as

$$y_{p,q} \approx H_{p,q} c_{p,q} + \eta_{p,q} \tag{4}$$

Where

$$c_{p,q} = d_{p,q} + \underbrace{j \sum_{(m,n)\in\Omega_{p,q}} d_{m,n} \langle g\rangle_{m,n}^{p,q}}_{u_{p,q}} = d_{p,q} + ju_{p,q} \tag{5}$$

$$H_{k,l}^{(c)} = \iint H(\tau,\nu) e^{j\pi(2l\tau_0\nu - 2k\nu_0 + \tau\nu)} d\nu d\tau \tag{6}$$

Represents the channel coefficients at q[th] symbol p[th] subcarrier frequency, and

$$d_{k,l}^{(c)} = \sum_{p,q} a_{k+p,l+q} j^{p+q+p(q+2l)} A_g^*(q\tau_0, pv_0) \tag{7}$$

is the virtual transmitted symbol at (p , q), with

$$u_{p,q} = \sum_{(m,n)\in\Omega_{p,q}} d_{m,n} \langle g \rangle_{m,n}^{p,q} \tag{8}$$

being the imaginary part of the interference from the neigboring FT points. When known pilots are transmitted at that FT point and its neighbourhood $\Omega_{p,q}$, the quantity in (5) can be approximated. This can be serve as a pseudo pilot [9] to compute an estimate of the CFR at the corresponding FT point, as, for example,

$$\hat{H}_{p,q} = \frac{y_{p,q}}{c_{p,q}} \approx H_{p,q} + \frac{\eta_{p,q}}{c_{p,q}} \tag{9}$$

This observation underlies the IAM schemes to be described in detail in the sequel.

III. IAM PREAMBLES WITH GUARD SYMBOLS

From equation (9) that the preamble should be so organised as to result in pseudo-pilots of maximum magnitude [9]. Therefore, the training symbols surrounding the pilot $d_{p,q}$ should be such that all terms in (8) have the same sign so that they add together. Moreover, this should occur for all frequencies p.

To this end, one needs to know the interference weights $\langle g \rangle_{m,n}^{p,q}$ for the neighbors $(m,n)\in\Omega_{p,q}$ of each FT point (p,q) of interest. These can be a priori computed based on the prototype filter g employed. Furthermore, it can be shown that, for any choice of g, these weights follow a specific pattern, which, for the definition of φ_0 adopted here, and for all q, can be written as

$$\begin{matrix} (-1)^p\delta & -\beta & (-1)^p\delta \\ -(-1)^p\gamma & d_{p,q} & (-1)^p\gamma \\ (-1)^p\delta & \beta & (-1)^p\delta \end{matrix} \tag{10}$$

With the horizontal direction corresponding to time and the vertical one to frequency. The above quantities can be given by

$$\beta = e^{-j\frac{2\pi}{M}\frac{L_g-1}{2}} \sum_{l=0}^{L_g-1} g^2(l) e^{j\frac{2\pi}{M}l} \tag{11}$$

$$\gamma = \sum_{l=\frac{M}{2}}^{L_g-1} g(l) g(l-\frac{M}{2}) \tag{12}$$

$$\delta = -je^{-j\frac{2\pi}{M}\frac{L_g-1}{2}} \sum_{l=M/2}^{L_g-1} g(l) g(l-\frac{M}{2}) e^{j\frac{2\pi}{M}l} \tag{13}$$

and are positive and, clearly, smaller than one. Generally, $\beta, \gamma > \delta$. for example, $\gamma = 0.5644$ $\beta = 0.2393$ and $\delta = 0.2058$. for the g's employed in part V. This observation, along with the symmetry of the above pattern, will be seen to underly the development of the IAM method proposed in Section IV.

A. IAM-R

In order to simplify the task of generating pseudo-pilots of large magnitude, it was proposed in [15, 14] to place nulls at the first and third FBMC/OQAM symbols of the preamble, that is, $a_{p,0} = a_{p,2} = 0, p = 0,1,...., M-1$. Then, the imaginary part of the pseudo-pilot at (p,1) will only come from the symbols at the positions $(p \pm 1,1)$. Obviously, these pilots should be OQAM symbols of maximum modulus, d. Moreover, in view of the pattern (10), they should satisfy the condition $d_{p+1,1} = -d_{p-1,1}$ for all

p, in order to yield pseudo pilots of maximum magnitude, namely $\left|d_{p,1} + j2\beta d_{p+1,1}\right| = d\sqrt{1+4\beta^2}$. an example for M =8 and OQPSK modulation is shown in Fig 1(a). This method is called IAM-R to signify the fact that its pilot symbols are real.

B. IAM-C

Regardless, it can be verified from (5) that one can do better in maximizing the magnitude of the pseudo-pilots if imaginary pilot symbols $jd_{p,1}$ of maximum modulus d are also allowed in the preamble. The reason is that the corresponding pseudo-pilot would then be imaginary and hence of larger magnitude than in IAM-R, namely $\left|d_{p,1} + u_{p,1}\right|$, provided that the signs of the pilot symbol and its neighbors are properly chosen so that $d_{p,1}u_{p,1} > 0$. Such an IAM scheme was first suggested in [11], and was called IAM-I to signify the presence of imaginary pilots. The middle preamble vector comprises of triplets, with each of them following the above principle and the pilots in each triplet are otherwise choosen in random and independently of the other triplets. Hence, imaginary pseudo-pilots result in only one third of the subcarriers, whereas the rest of them deliver complex pseudo-pilots of smaller magnitude, $d\left|(1+\beta) + j\beta\right|$. With a minor but significant modification, one can improve upon this and obtain pseudo-pilots that are either purely real or imaginary at all the subcarriers. The idea is to simply set the middle FBMC/OQAM symbol equal to that in IAM-R but with the pilots at the odd subcarriers multiplied by j, as shown in [8] and independently in [6]. All of the resulting pseudo-pilots are then of maximum magnitude, $d\left|1+2\beta\right|$. Again, for the M = 8 OQPSK example, this will result in the preamble shown in Fig. 1(b). This scheme was called IAM-C. Note that this preamble is, firmly speaking, not OQAM. Regardless, it can still be fed to the synthesis filter bank and perfectly reconstructed at the analysis bank.

C. IAM-C EXTENDED(E IAM-C)

It was shown in [6] that the pseudo-pilots generated by IAM-C are of the maximum possible magnitude. However, this optimality is constrained to the 3-symbol preambles with nulls at their sides. The property mentioned at the end of part II suggests that an alternative

$$
\begin{array}{ccc}
\begin{array}{ccc}
0 & 1 & 0 \\
0 & -1 & 0 \\
0 & -1 & 0 \\
0 & 1 & 0 \\
0 & 1 & 0 \\
0 & -1 & 0 \\
0 & -1 & 0 \\
0 & 1 & 0
\end{array}
&
\begin{array}{ccc}
0 & 1 & 0 \\
0 & -j & 0 \\
0 & -1 & 0 \\
0 & j & 0 \\
0 & 1 & 0 \\
0 & -j & 0 \\
0 & -1 & 0 \\
0 & j & 0
\end{array}
&
\begin{array}{ccc}
j & 1 & -j \\
-1 & -j & 1 \\
-j & -1 & j \\
1 & j & -1 \\
j & 1 & -j \\
-1 & -j & 1 \\
-j & -1 & j \\
1 & j & -1
\end{array}
\\
\text{(a)} & \text{(b)} & \text{(c)}
\end{array}
$$

Fig 1: preamble structures for (a) IAM-R (b) IAM-C and (c) E IAM-C

Preamble structure which employs the side symbols as well could yield pseudo pilots that are stronger than those of IAM-C. It turns out that such a preamble can really be formed in an analogous way as previously, by using the left- and right-hand sides of the neighborhood (10) as well. Thus, at an odd-indexed subcarrier p, with the pilot dj in the middle, one should place $\mp d$ at the right hand side, and its negative, $\pm d$ at the left hand side. These will then contribute $\pm d\gamma j$ each to the interference. In an analogous way, at an even-indexed subcarrier p, the middle pilot, say $\pm d$, will get a total interference of $\pm dj$ from its (p,0) and (p,2) neighbours, if these are chosen as $\pm dj$ and $\mp dj$, respectively. It can then be verified that, because of (10), the interference components from the corner neighbors cancel each other, and hence have no contribution to the pseudo-pilot. However, this loss is not significant since d is significantly smaller than g. The resulting pseudo-pilots are again either purely real or imaginary, with magnitude $d\left|1+2(\beta+\gamma)\right|$, which is clearly larger than that achieved by IAM-C.

IV. OPTIMAL IAM-C(O IAM-C)

Inspired by IAM-C and observing that with the triplet using $d_{k\pm1}^{(c)} \approx (1+\delta) \pm j\delta$ have smaller power compared with $d_{k,l}^{(c)}$, one may think of finding the optimal complex valued $d_{m,l}$, carefully so that the demodulated symbols $d_{k,l}^{(c)}$ will have the maximum power.

To further increase the power of demodulated preamble symbols $d_{k,l}^{(c)}$, the two neighbouring columns can further more be utilized by assigning proper complex valued symbols rather than zeros. For convenience, the following notations are used to simplify our derivation.

$$A_g(0,\pm v_0) = \delta, \quad A_g(\pm\tau_0, 0) = \beta, \quad A_g(\pm\tau_0, \pm v_0) = \gamma, \quad A_g(\pm\tau_0, \pm 2v_0) = \eta$$

For distinct pulse shapes g(t) the above parameters are distinct in general. For well designed pulse shapes, $A_g(n\tau_0, mv_0)$ attenuates very fast with increasing m, n. For the demodulated symbol $d_{k,l}^{(c)}$, a weighting matrix W can be formulated to calculated prominent interference elements from neighbouring symbols. As shown in (7), the sign of the elements in w will be distinct for distinct value of l . Therefore two different realisations are listed as w_{leven} when l is an even number and w_{lodd} .

$$w_{leven} = \begin{bmatrix} -j\eta & 0 & j\eta \\ -j\gamma & -j\delta & j\gamma \\ -j\beta & 1 & j\beta \\ -j\gamma & j\delta & j\gamma \\ -j\eta & 0 & j\eta \end{bmatrix}$$

w_{leven} and w_{lodd} only differs from each other by altering the sign of all the elements in the second row and the fourth row. Since the first and the third column of the weighting matrix w also furthermore contribute to intrinsic interference, more power will be conveyed to $d_{k,l}^{(c)}$ by assigning proper complex valued symbols rather than zeros. The symmetric and fast decay nature of the weighting matrix w in the central column are repetition of a quadruplet, the symbols in the other two columns should also be grouped into quadruplet. As shown in Appendix A, given the weighting matrix w, the optimal preamble sequence is as shown in Fig 2

$$\begin{matrix} j & 1 & j \\ -j & -j & j \\ j & -1 & -j \\ -j & j & j \end{matrix} \qquad \begin{matrix} j & 1 & j \\ 1 & j & 1 \\ j & -1 & j \\ 1 & -j & 1 \end{matrix}$$

(a) (b)

Fig 2: optimal preamble structures for (a) optimal-IAM (l_{even}) (b) optimal-IAM (l_{odd})

This new method was named as optimal-IAM and the preamble structures are shown in Fig 2, and the preamble structure appears repeatably to formulate the preamble sequence.

V. SIMULATION RESULTS

where N =512, 1024, 2048 FFT/IFFT was used with a convolutional channel coding at rate 1/2, we simply use a relatively small FFT/IFFT .The first three IAM preambles, namely IAM-R, IAM-I and E- IAM-C, have the same energy on preamble sequences and share a similar structure (0 X 0) and there will be frequently

compared with each other in our following discussion. The last IAM preambles, namely Optimal- IAM-C, have the same structure as defined by O-IAM-C (l_{even}) and O-IAM-C (l_{odd}) but three times the energy of the other three preambles. The normalized MSE $\frac{\|H - \hat{H}\|^2}{\|H\|^2}$ where H is the channel frequency response vector & $\hat{H}$ its estimate, is plotted with respect to the E_b / N_0 (SNR). BER simulation versus E_b / N_0 (SNR), with E_b the useful bit energy and N_0 the mono-lateral noise density, has been carried out for different preamble sequence, as shown in Fig.1&2. The O-IAM-C method can always outperform E-IAM-C by at least 1.24 dB. Fig. 4 shows the MSE performance of the methods under study for M = 512, 1024 & 2048 subcarriers with an overlapping factor of K = 4 are considered in Figs. 4(a),(b) and (c), respectively. Observe that in three cases O-IAM-C outperforms the other schemes in the whole SNR range considered. Fig. 3 shows the BER performance of the methods under study for M=512, 1024 & 2048 subcarriers with an overlapping factor of K=4 are considered in Figs. 3(a), (b) and (c), respectively. Observed that in all the cases O-IAM-C outperforms E-IAM-C by at least 1.24 dB gain.

VI. CONCLUSION

The problem of preamble-based CE in FBMC/OQAM systems was revisited in this paper, with a aim on the interference approximation idea. The three most well renowned and efficient IAM variants, IAM-R, IAM-C and E-IAM-C, were reconsidered along with their associated preamble structures. Emphasis was put on IAM-C and its optimality within the class of 3- symbol preambles with all zeros side FBMC symbols. The possible gains from resting the last mentioned constraints of all zeros at the sides were investigated, on the basis of fully exploiting the symmetries in the interference weights of the first-order FT neighbours. An innovative preamble structure of this kind, optimal-IAM-C, was developed as an extension of the E- IAM-C idea and shown to result in pseudo-pilots of significantly larger magnitude. The superiority of O-IAM-C was demonstrated via results of simulation in a realistic (including interference from data) and fair scenario.

APPENDIX A

A. Derivation of the optimal IAM preamble structure

For odd l, given the preamble matrix

$$A_{l odd} = \begin{bmatrix} x_1 & 1 & x_5 \\ x_2 & j & x_6 \\ x_3 & -1 & x_7 \\ x_4 & -j & x_8 \end{bmatrix} \tag{15}$$

Where $|x_i| \le 1, i = 1,.....,8$ are complex valued variables, the demodulated symbols at corresponding points are as follows:

$$\begin{aligned} a_1^{(c)} &= 1 + 2\delta + j\underbrace{[\beta(x_5 - x_1) + \gamma(x_2 + x_4 + x_6 + x_8) + 2\eta(x_7 - x_3)]}_{y_1} \\ a_j^{(c)} &= j[1 + 2\delta + \underbrace{\beta(x_6 - x_2) + \gamma(x_1 + x_3 + x_5 + x_7) + 2\eta(x_8 - x_4)}_{y_2}] \\ a_{-1}^{(c)} &= -1 - 2\delta + j\underbrace{[\beta(x_7 - x_3) + \gamma(x_2 + x_4 + x_6 + x_8) + 2\eta(x_5 - x_1)]}_{y_3} \\ a_{-j}^{(c)} &= j[-1 - 2\delta + \beta\underbrace{(x_8 - x_4) + \gamma(x_1 + x_3 + x_5 + x_7) + 2\eta(x_6 - x_2)}_{y_4}] \end{aligned} \tag{16}$$

Then the preamble design problem becomes a optimisation process which try to find a realisation of $x_i, i = 1,...., 8$ that can maximise the minimum power of the demodulated symbols, i.e., that can maximise the minimum power of the demodulated symbols, i.e,

$$\{x_i, i = 1,..., 8\} = \arg \max_{x_i, i=,...,8} \min\{\left|a_1^{(c)}\right|^2, \left|a_j^{(c)}\right|^2, \left|a_{-1}^{(c)}\right|^2, \left|a_{-j}^{(c)}\right|^2\} \quad (17)$$

Under the constraint that $\left|x_i\right| \leq 1, i = 1,....., 8$.

In order to maximise the power, all the elements in each $a^{(c)}$ in (16) should be coherently added, that is,

$$phase(x_5 - x_1) = phase(x_2 + x_4 + x_6 + x_8) = phase(x_7 - x_3)$$

$$phase(x_6 - x_2) = phase(x_1 + x_3 + x_5 + x_7) = phase(x_8 - x_4) \quad (18)$$

Given (16), it is easy to figure out that the condition that maximise $\left|a_1^{(c)}\right|^2$ (e.g. when $jy_1 > 0$) will minimise $\left|a_{-1}^{(c)}\right|^2$, and vice versa. The same situation happens for $\left|a_j^{(c)}\right|^2$ and $\left|a_{-j}^{(c)}\right|^2$. Therefore the optimal solution to maximise the minimum power is to set y_1, y_3 purely real valued and y_2, y_4 purely imaginary, that is,

$$phase(y_1) = phase(y_3) = 0 \ \ or \ \pi$$

$$phase(y_2) = phase(y_4) = \pi / 2 \ \ or \ -\pi / 2$$

Therefore following equations can be formulated as

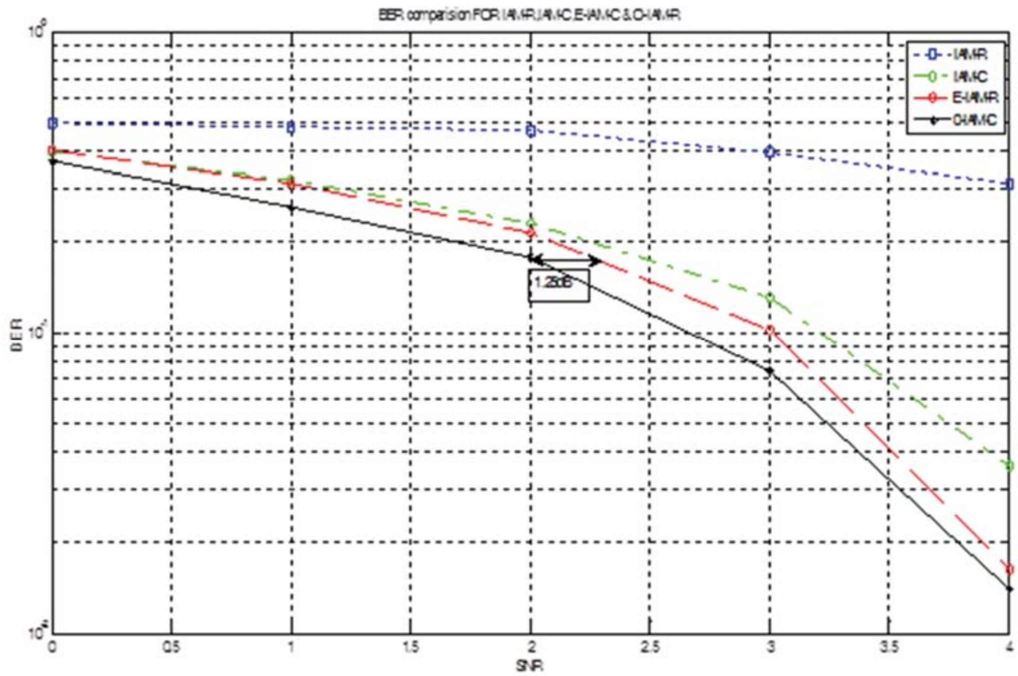
$$\begin{cases} \{\Im(x_5) = \Im(x_1), \Im(x_7) = \Im(x_3) & y_1, y_3 \quad \textit{purely real} \\ \Re(x_5)\Re(x_1) < 0, \Re(x_7)\Re(x_3) < 0 & \textit{coherent addition} \\ \Re(x_1 + x_3 + x_5 + x_7) = 0 & \\ \Re(x_6) = \Re(x_2), \Re(x_8) = \Re(x_4) & y_2, y_4 \quad \textit{purely imaginary} \\ \Im(x_6)\Im(x_2) < 0, \Im(x_8)\Im(x_4) < 0 & \textit{coherent addition} \\ \Im(x_2 + x_4 + x_6 + x_8) = 0 & \end{cases} \quad (19)$$

Where $\Re(.)$ denotes the real part operator and $\Im(.)$ denotes the imaginary part operator. It is clear from (17), (18) and (19) that the power of $y_i, i = 1,...., 4$ can reach the maximum only if $\left|x_i\right| = 1, i = 1,....., 8$, i.e., each element in the preamble utilises the maximum power available. Hence (5.21) can be rewritten as

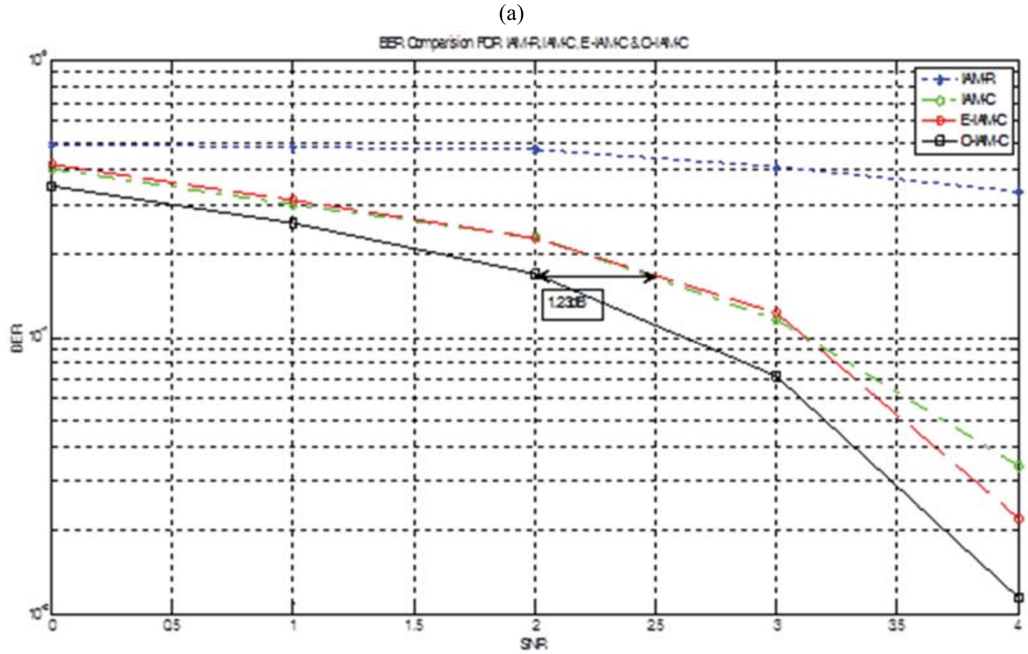
$$\begin{cases} \Im(x_5) = \Im(x_1), & \Re(x_5) = -\Re(x_1) \Rightarrow x_1 = -x_5^* \\ \Im(x_7) = \Im(x_3) & \Re(x_7) = -\Re(x_3) \Rightarrow x_3 = -x_7^* \\ \Re(x_6) = \Re(x_2), & \Im(x_6) = -\Im(x_2) \Rightarrow x_2 = x_6^* \\ \Re(x_8) = \Re(x_4), & \Im(x_8) = -\Im(x_4) \Rightarrow x_4 = x_8^* \end{cases} \quad (20)$$

Hence the preamble matrix A_{lodd} described in (15), we get is

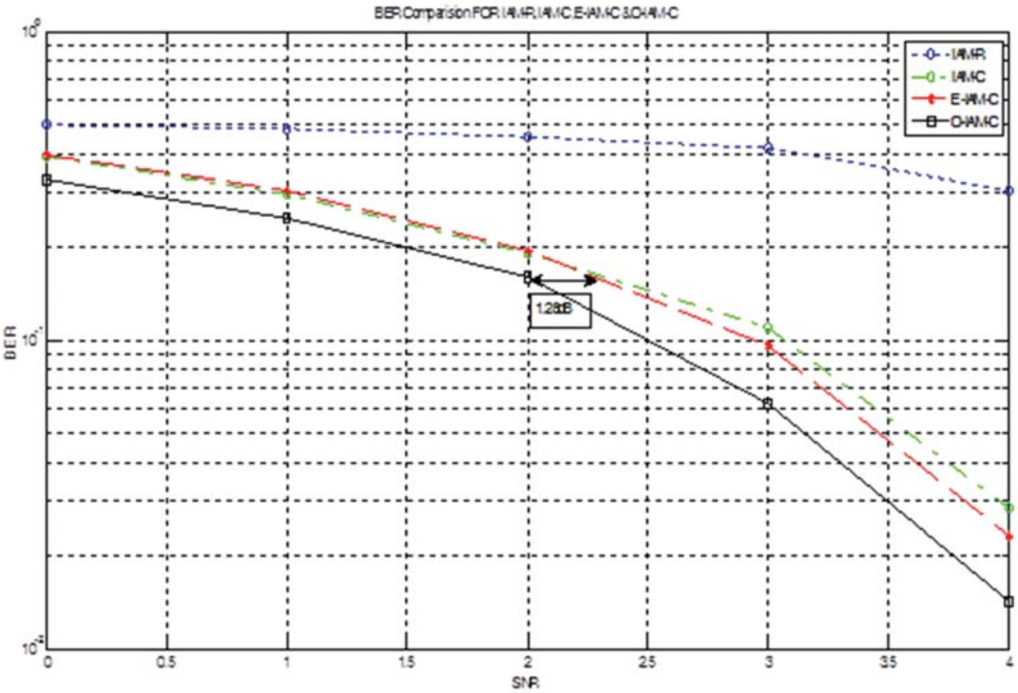
$$A_{lodd} = \begin{bmatrix} x^* & 1 & -x \\ -jx & -j & jx^* \\ x^* & -1 & -x^* \\ -jx & -j & x^* \end{bmatrix}$$


BER comparision FOR IAMR,IAMC,E-IAMC & C-IAMR
IAMR
IAMC
E-IAMR
C-IAMC
1.25dB
BER
SNR
0
0.5
1
1.5
2
2.5
3
3.5
4

(a)

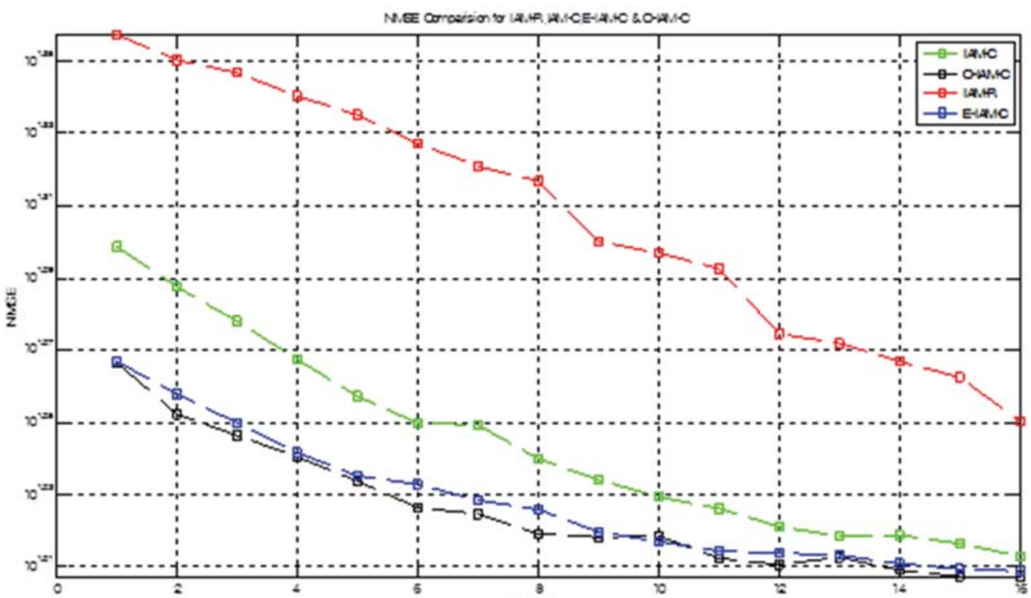

BER Comparision FOR IAM-R,IAMC,E-IAMC & C-IAMC
IAMR
IAMC
E-IAMC
C-IAM-C
1.23dB
BER
SNR
0
0.5
1
1.5
2
2.5
3
3.5
4

(b)

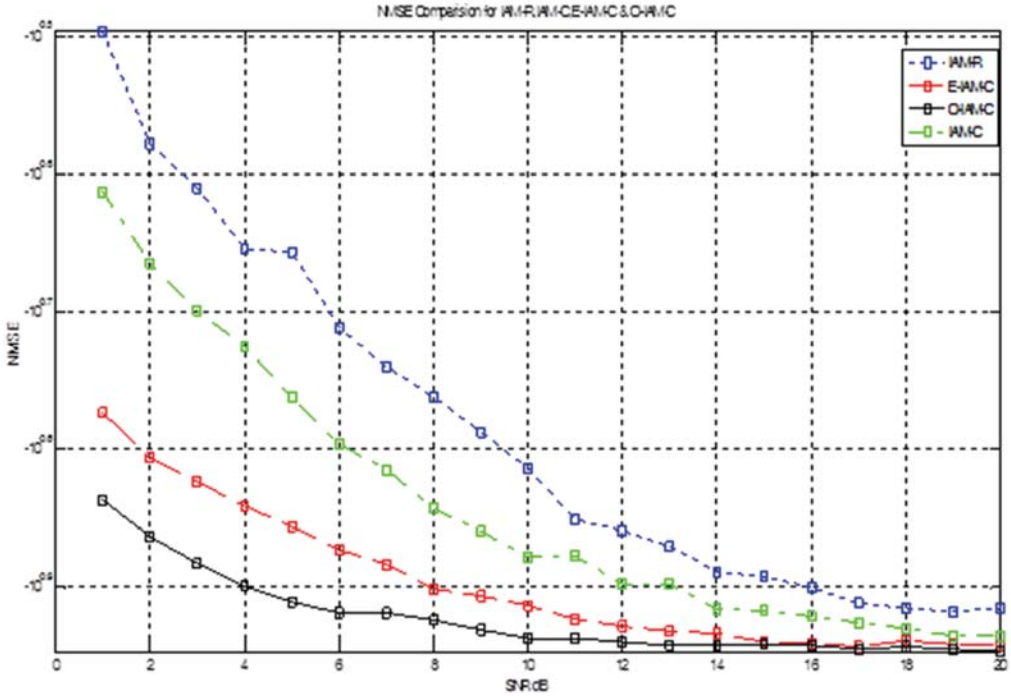


(c)

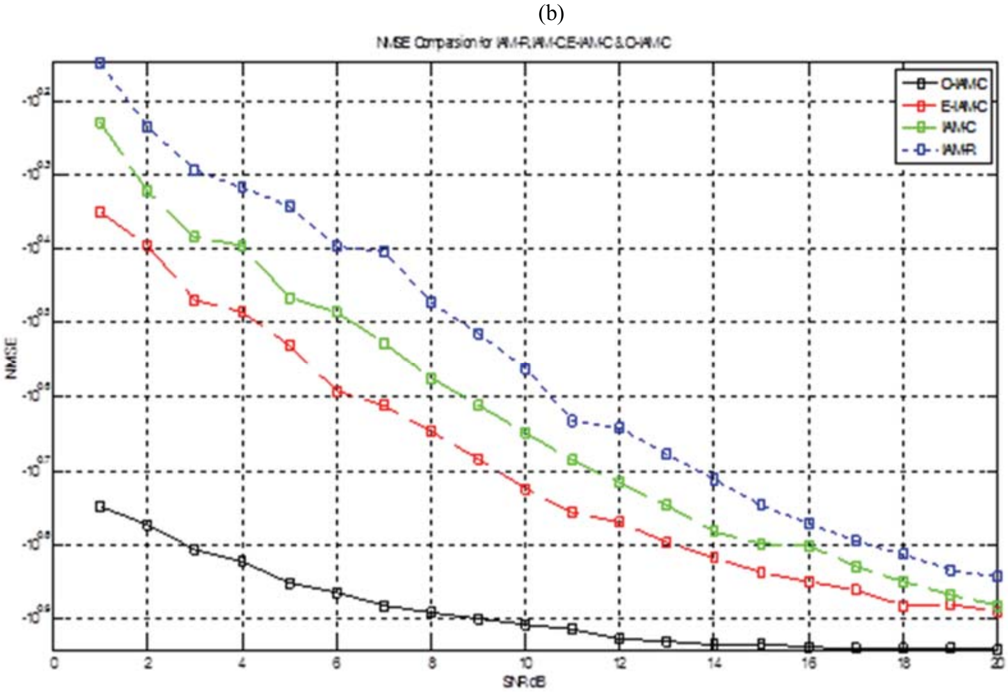
Fig3: BER vs. SNR with QPSK modulation .Rayleigh channels were assumed, with
(a) Filter banks with M=512,K=4(overlapping factor), (b) Filter banks with M=1024, K=4(overlapping factor) and (c) Filter banks with M=2048, K=4(overlapping factor)



(a)



(b)



(c)

Fig4: Estimation (NMSE) performance of preamble based methods .Rayleigh channels were assumed, with Filter banks with M=512, K=4(overlapping factor), (b) Filter banks with M=1024, K=4(overlapping factor) and (c) Filter banks with M=2048, K=4(overlapping factor)

REFERENCES

[1] P. Amini, R. Kempter, and B. Farhang-Boroujeny, “A comparison of alternative filterbank multicarrier methods for cognitive radio systems,” Proc. SDR'06.

[2] J. G. Andrews, A. Ghosh, and R. Muhamed, Fundamentals of WiMAX: Understanding Broadband Wireless Networking, Prentice-Hall, 2007.

[3] F. Bader and M. Shaat, “Pilot pattern adaptation and channel estimation in MIMO WiMAX-like FBMC system,” Proc. ICWMC'10.

[4] M. Bellanger et al., “OFDM and FBMC transmission techniques: a compatible high performance proposal for broadband powerline communications,” Proc. ISPLC' 10.

[5] N. Benvenuto et al., “Analysis of channel noise in orthogonally multiplexed OQAM signals,” Proc. GLOBECOM'93.

[6] J. Du and S. Signell, “Novel preamble-based channel estimation for OFDM/OQAM systems,” Proc. ICC'09.

[7] J.-P. Javaudin, D. Lacroix, and A. Rouxel, “Pilot-aided channel estimation for OFDM/OQAM,” Proc. VTC'03 (spring).

[8] B. Le Floch, M. Alard, and C. Berrou, “Coded orthogonal division multiplex,” Proc. IEEE, June 1995.

[9] C. Lele et al, “Channel estimation methods for preamble-based OFDM/OQAM modulations,” European Trans. Telecommunications, pp. 741.750, 2008.

[10] C. Lele, R. Legouable, and P. Siohan, “Channel estimation with scattered pilots in OFDM/OQAM,” Proc. SPAWC'08.

[11] C. Lele, P. Siohan, and R. Legouable, “2 dB better than CP-OFDM with OFDM/OQAM for preamble-based channel estimation,” Proc. ICC'08.

[12] P. Siohan, C. Siclet, and N. Lacaille, “Analysis and design of OFDM/OQAM systems based on filter bank theory,” IEEE Trans. Signal Processing, May 2002.

[13] R. van Nee and R. Prasad, OFDM for Wireless Multimedia Communications, Artech House Publ., 2000.

[14] P. Siohan, C. Siclet and N. Lacaille, “Analysis and design of OFDM/OQAM. systems based on filterbank theory”, IEEE Transactions on Signal Processing, vol. 50, no. 5, pp. 1170–1183, May 2002.

[15] H. Bolcskei, “Orthogonal Frequency Division Multiplexing Based on Offset-QAM”, Advances in Gabor Analysis, Birkhauser, 2003.

[16] Deliverable D4.1: “MIMO channel estimation and tracking,” PHYDYAS document (online) http://www.ictphydyas org/delivrables/PHYDYAS-D4.1.pdf/view.

[17] E.Koffidis and D.Katselis, “Improved Interference Approximation Method For Preamble-Based Channel Estimation in FBMC/OQAM,” Proc.EU-SIPCO’11, Barcelona, Spain, 29 Aug.-2 Sept.2011.

[18] C. Lele, OFDM/OQAM Modulation: Channel Estimation Methods, and Applications to Multicarrier CDMA and Multi-antenna Transmission, Ph.D. thesis, CNAM, 2008.